BBC RD 1976/20

BBC

RESEARCH DEPARTMENT

1974 THE QUEEN'S AWARD TO INDUSTRY

REPORT

An experimental 4-phase differential-phase-shift-keying system to carry two high-quality digital sound signals

M.J. Kallaway, M.A.

Research Department, Engineering Division
THE BRITISH BROADCASTING CORPORATION August 1976

BBC RD 1976/20
UDC 621.376.4:
534.86

AN EXPERIMENTAL 4-PHASE DIFFERENTIAL-PHASE-SHIFT-KEYING SYSTEM TO CARRY TWO HIGH-QUALITY DIGITAL SOUND SIGNALS

M.J. Kallaway, M.A.

Summary

The report describes an experimental 4-phase differential-phase-shift-keying system to carry two high-quality sound signals. The system has possible applications in outside broadcast links and satellite broadcasting.

The system offers a good compromise between the requirements of carrier power, signal bandwidth and instrumental simplicity. The factors that led to the choice of this particular type of system are discussed in some detail.

The report concludes that the performance of the experimental system in the presence of noise is within about 0·5 dB of the theoretical limit. Measurements have been made to determine the performance of individual units within the system. The most critical areas of instrumentation are identified and the effect of instrumental imperfections within these critical areas is analysed. Before recommendations can be made, further tests of the system, particularly under multipath propagation conditions, are required.

Issued under the authority of

G.D. Monteath

Head of Research Department

Research Department, Engineering Division,
BRITISH BROADCASTING CORPORATION
August 1976
(EL-121)

AN EXPERIMENTAL 4-PHASE DIFFERENTIAL-PHASE-SHIFT-KEYING SYSTEM TO CARRY TWO HIGH-QUALITY DIGITAL SOUND SIGNALS

AN EXPERIMENTAL 4-PHASE DIFFERENTIAL-PHASE-SHIFT-KEYING SYSTEM TO CARRY TWO HIGH-QUALITY DIGITAL SOUND SIGNALS

M.J. Kallaway, M.A.

1. Introduction

Increasing use is being made of digital techniques for the distribution and broadcasting of high-quality sound signals. The coding of an analogue signal into p.c.m. form greatly increases the immunity of the signal to transmission impairments. As long as the individual digits of the signal can be recognised by the receiver without error, the original audio signal can be perfectly reconstructed subject only to the original quantising errors. Digital distribution networks are effectively transparent whereas, when a signal is distributed in analogue form, the distribution network is bound to degrade the quality of the signal to some extent.

At present, high-quality sound signals are distributed from the BBC's London studios to the principal v.h.f./f.m. transmitters in digital form. Thirteen individual audio p.c.m. channels are multiplexed together to form a composite bit-stream with a bit-rate of 6·336 Mbit/s. The bearer circuits are s.h.f./f.m. links primarily designed for the distribution of monochrome 625-line television signals. Moreover, except where off-air pick-up is employed, sound signals have for a number of years been distributed in digital form to all television transmitters, using the sound-in-syncs system. Here, the digital information is carried within the line synchronising pulses of the video signal.

One area where digital techniques could also be applied is in contribution links, in particular outside broadcast (OB) radio links. In the existing equipment, the audio-signal frequency-modulates a v.h.f. carrier. The system is found to be acceptable for monophonic signals but is not well suited for stereophonic applications, using the standard pilot-tone multiplex system, because the range of the equipment is greatly reduced in this mode. If a digital system were used, it might be possible to achieve stereo operation over a much greater range.

Primarily with this application in mind, an investigation was started on an efficient system for transmitting binary signals on a carrier, at a rate appropriate for two high-quality sound signals. The main object of the experiment, which included the construction of a modem, was to investigate in the laboratory the basic performance of the digital transmission system compared to f.m., under various conditions that might be met in practice. It is hoped that the knowledge gained from this work will be helpful in considering future applications of digital techniques to sound signal distribution and, possibly, in direct broadcasting from satellites.

The modulation system used in the experimental equipment was 4-phase differential-phase-shift-keying (d.p.s.k.). The equipment comprised a modulator and demodulator (modem) operating at an intermediate frequency of 10·7 MHz. The factors that governed the choice of signalling system are discussed first in the report. The instrumentation is then described and the performance of the experimental modem is assessed.

2. Basic system parameters

2.1. Bit-rate

Simple linear quantisation of an audio signal requires 13 bits per sample,[1] for acceptable broadcast quality. The usual sampling frequency is 32 kHz, leading to an audio bit-rate for a two-channel system of 832 kbit/s.

The number of bits per sample can be reduced to 10, using 'near-instantaneous' digital companding,[2] leading to a bit-rate of about 646 kbit/s. The groups or 'data words' of 10 bits that are transmitted are the more significant bits chosen from the original 13 bits so as to just accommodate the largest signal amplitude within a specified time-interval (32 samples). A two-bit scale-factor word is transmitted once for each time-interval to indicate the exact position of the 10 transmitted bits within the original 13-bit data words.

The two digital audio signals are next formed into a serial bit-stream and a framing pattern is added at this stage to enable the individual data words to be identified. The framing pattern typically increases the bit-rate by 1%,[3] leading to a total bit-rate of 652 kbit/s for a two-channel system.

The quoted bit-rate of 652 kbit/s does not include an allowance for any error-protecting bits except for a single parity bit in the scale-factor word. The form of errors in a 4-phase d.p.s.k. system is unusual (see Section 2.7) and any error-protecting or correcting scheme employed must take this into account. It was felt that it would be wise to explore the properties of the system fully before deciding on the type of error-protection to be used.

A system for reducing the effect of errors is likely to increase the audio bit-rate by 3 to 8%[3] and the experimental modem has been designed to accommodate this increase if necessary, with little modification.

There was no equipment available for combining two digital audio signals into a serial bit-stream suitable for work on the experimental modem described here. In all the experimental work described in this report, therefore, the audio signal has been simulated by a 652 kbit/s pseudo-random sequence.

2.2. Choice of signalling system

The digital signalling system should ideally be rugged,

simple to instrument and use as little bandwidth as possible. Unfortunately, a system tends to meet one objective at the expense of another objective and the final choice of signalling system must be a compromise. The arguments that governed the choice of signalling system used in the experimental modem are briefly summarised below. A much fuller treatment of the properties of digital signalling systems is given in Reference 4.

An r.f. carrier can be quantised in either amplitude or phase, or in both. The transmitted signal is then made up from a set of symbols each with N possible states. Each symbol carries $\log_2 N$ bits of information and hence the transmitted bit-rate equals $f_s \log_2 N$ where f_s is the symbol rate.

The minimum theoretical bandwidth required to transmit such a signal is f_s, if the form of modulation is double-sideband. For practical reasons which are discussed later, a greater bandwidth is required, a more realistic figure being about $1{\cdot}3\ f_s$. If the number of discrete states N of each symbol is increased then, for a given bit-rate, the bandwidth of the signal is reduced.

The carrier-to-noise ratio required for a given error-rate is governed by the number of signalling states. An error will occur whenever a noise peak carries the signal over half the difference between adjacent signal states. If the number of signal states is increased, the states move closer together and the carrier power for a given error-rate must be increased.

The relationship between carrier power, bandwidth and number of signalling states has been discussed elsewhere.[4,5] If N has a value greater than about four, any increase in N designed to save bandwidth will require a disproportionate increase in carrier power. This trend towards a law of diminishing returns is common to all digital signalling systems. Unless bandwidth is at a real premium, the benefits gained by using a system with more than eight states are very small. Moreover, instrumentation of systems with large numbers of states is very difficult if maximum performance is to be achieved.

When the number of states is small, a signalling system that quantises the phase dimension only is very economical in peak carrier power. In phase modulation, the amplitude of the signal is constant at each of the phase states. However, in the form of phase modulation used here, the transitions between phase states are not made at constant carrier amplitude. If the carrier-amplitude were constant during transitions the signal would have sidebands similar to those of an f.m. signal, and the bandwidth required for undistorted transmission would be much greater. It is convenient to regard the signal as being produced by amplitude-modulating two carriers in phase-quadrature, so that although the phase states are at constant amplitude the transitions are not. This type of system is usually termed phase-shift-keying (p.s.k.) although the properties of the signal have more in common with those of amplitude modulation than true phase-modulation.

After careful consideration of the above arguments, a 4-phase p.s.k. system was chosen as being most suitable for the experimental modem. The system provides a good compromise between the requirements of carrier power, bandwidth and instrumental simplicity. Because of these properties, the system is favoured for many digital signalling application in other fields.

2.3. Differential coding and synchronous demodulation

The changing phase pattern of the p.s.k. signal can be detected by comparing the received signal phase with a reference phase, that must bear a fixed relationship with the original unmodulated carrier for the message to be correctly interpreted. The p.s.k. signal, for a random message, has equal probabilities of being in any one of the four phases. Such a signal carries no information about the phase of the original carrier and so it is impossible to derive the correct reference phase for demodulation. A reference phase can be recovered if extra information is added to the p.s.k. signal. The extra information can take a variety of forms, the most obvious being to add a residual carrier component. Another possibility is to add redundancy to the data signal so that incorrect phasing can be recognised and a suitable correction made. The extra information is carried at the expense of an increase in carrier power or an increase in signal bandwidth, neither of which is desirable.

The problem of establishing a reference phase can be avoided if the p.s.k. signal is differentially coded; the process is illustrated in Fig. 1. The transmitted message is coded into phase-changes between one symbol of the differential-phase-shift-keyed (d.p.s.k.) signal and the next. The data signal to be transmitted is split up into pairs of digits, the four possible values of which are 00, 01, 10, and 11. Each pair of digits is assigned a specific phase-change, as shown in Fig. 1(a). Fig. 1(b) shows the "quadrants" of the d.p.s.k. signal and Fig. 1(c) shows, for an arbitrary bit-stream, the transmitted phase-changes and the instantaneous quadrant of the d.p.s.k. signal. The d.p.s.k. signal is initially assumed to be in quadrant 1.

A d.p.s.k. signal can be detected simply by comparing the phase of the present symbol with that of the previous symbol. For example, a demodulator can employ a phase detector, one of whose inputs is fed with a direct signal while the other input is fed with a signal delayed by one symbol period. The output from the phase detector indicates the phase difference between successive symbols and can be decoded to give the transmitted message.

In another method, the 4-phase d.p.s.k. signal is synchronously demodulated by the in-phase and quadrature components of a reference carrier whose phase is locked to one of the four phase-states. The outputs from the two synchronous demodulators indicate the relative phase between the reference and the d.p.s.k. signal. By sampling these two outputs the quadrant occupied by the d.p.s.k. signal can be determined and stored. Successive comparisons between the stored values enable the phase-change to be found and the message decoded. For successful operation the phase of the reference may have any one

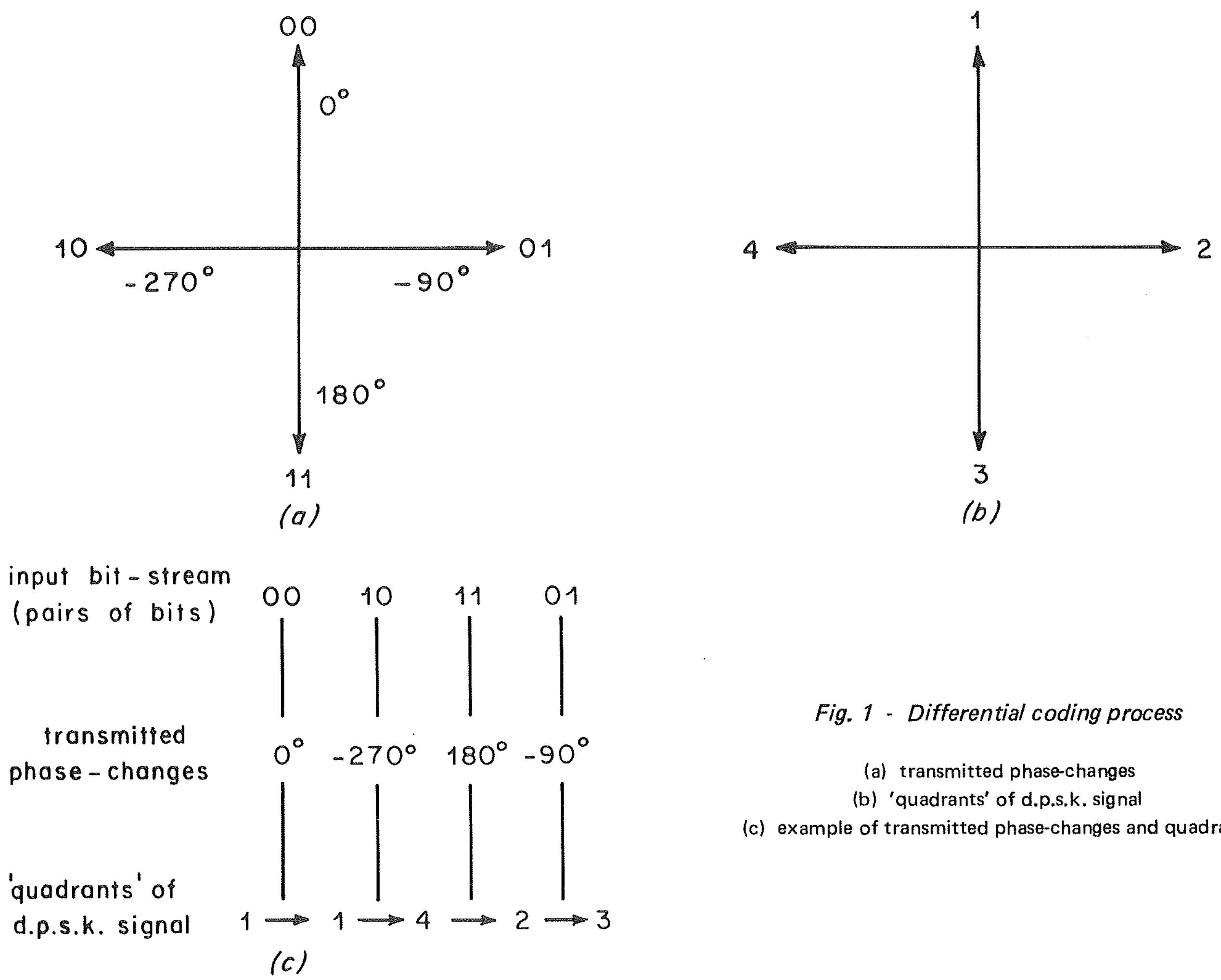


Fig. 1 - Differential coding process

(a) transmitted phase-changes
(b) 'quadrants' of d.p.s.k. signal
(c) example of transmitted phase-changes and quadrants

of four values. Methods of establishing such a reference are described in Section 2.5.

The synchronous demodulation method is preferable and was used in the experimental modem for the following reasons. In the first method, both signals applied to the phase detector contain noise, so that the effective signal-to-noise ratio at the output of the phase detector is lower than at the input. To compensate for this loss of signal-to-noise ratio, the carrier power must be increased by about 2·5 dB. If the d.p.s.k. signal is synchronously demodulated, noise can be filtered from the reference carrier and there is negligible loss of signal-to-noise ratio during the demodulation process.

In the above discussion we have considered only one of many ways of avoiding the need for a fixed carrier reference. There is another widely used method in which the signals derived from two synchronous demodulators can provide two independent bit-streams. At first sight, such a method may appear attractive in the case when the p.s.k. signal is to carry two independent audio signals. Whilst it is true that a form of differential modulation can be used on each axis the two axes cannot be identified one from another without the addition of extra information. Consequently, this method offers no real advantage over the method that has been used.

2.4. Spectrum shaping and intersymbol interference

Ideally, at the receiver, each symbol of the message should be identified without interference from the other symbols of the message. A digital signal has several discrete states; the greater their separation, the easier it is to distinguish between them in the presence of noise. Intersymbol interference will cause each of the states to occupy a range of values at the receiver, reducing the effective separation of the states and lowering the noise margin of the signal.

The amount of intersymbol interference on the received signal is set by the spectrum shaping, intentional and unintentional, applied to the signal throughout the system. For a carrier system, the spectrum shaping can be applied to the signal before modulation, after demodulation and also between the modulator's output and the demodulator's input. From the viewpoint of intersymbol interference, the distribution of the spectrum shaping is unimportant: only the overall characteristic matters. However, when considering the effects of noise, the distribution assumes greater importance.

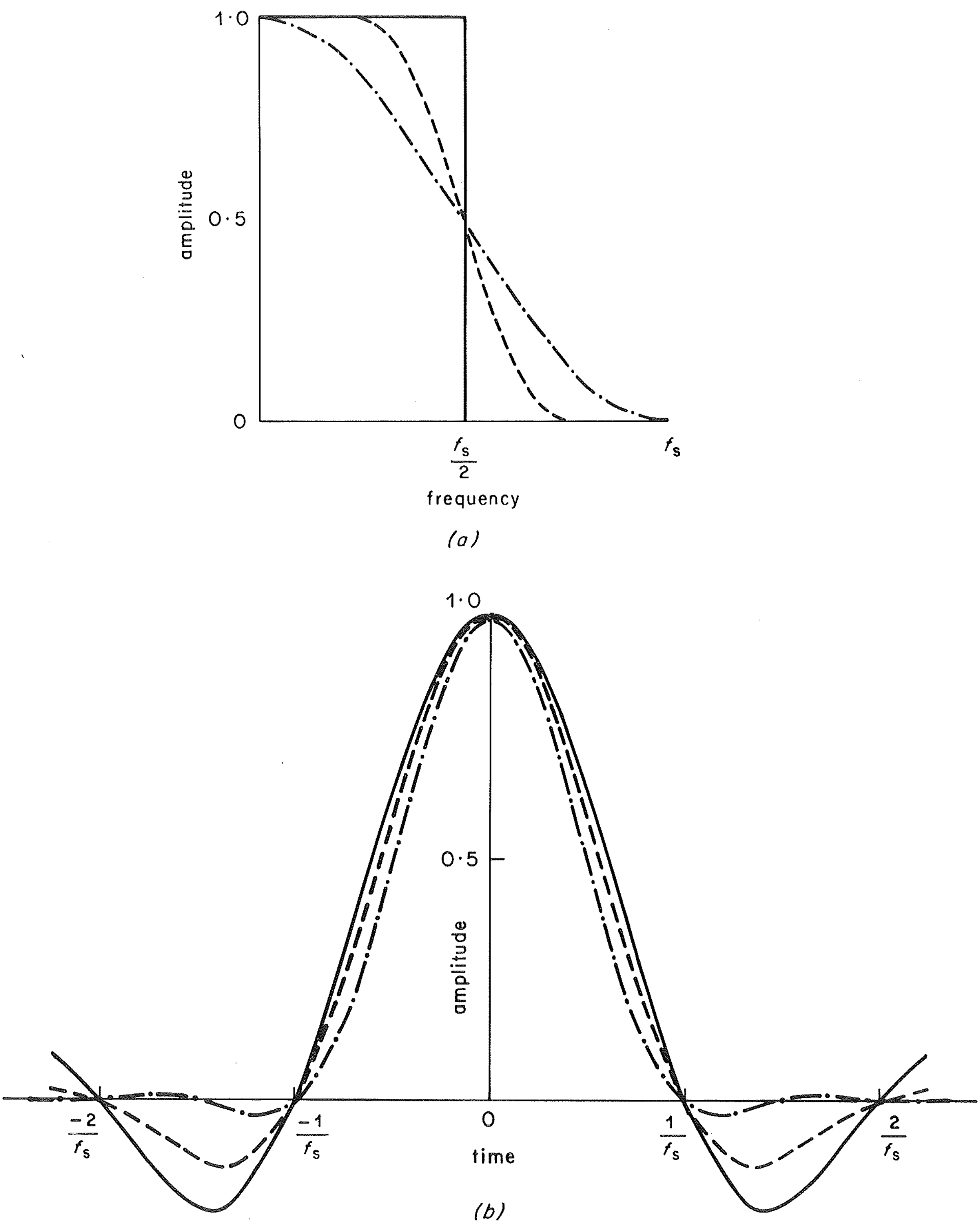


Fig. 2 - Amplitude and impulse responses of cosine roll-off filters

———— = 0% cosine roll-off, – – – – – = 50% cosine roll-off, –·–·–·= 100% cosine roll-off

(a) amplitude responses, percentage cosine roll-off = $\left[\dfrac{\text{upper cut-off frequency} - \frac{f_s}{2}}{\frac{f_s}{2}}\right] \times 100\%$

(b) impulse responses

It is convenient to transform the individual characteristics of all the spectrum shaping filters into a single low-pass characteristic between the input to the modulator and the output of the demodulator. The properties of the carrier system are then a simple extension of the properties of the baseband system. This transformation can be made for a double-sideband carrier system providing the r.f. spectrum shaping is symmetrical about the carrier frequency. In the case of a p.s.k. or d.p.s.k. system, it has been assumed earlier that the signal is produced by amplitude modulation of two carriers in phase-quadrature. Any asymmetry of amplitude or group-delay response about the carrier frequency will produce crosstalk between the two quadrature axes of modulation. This crosstalk cannot be corrected by baseband filtering of the signals prior to modulation or after demodulation.

A baseband data signal which is a train of impulses with symbol rate f_s can be passed without intersymbol interference through a low-pass filter whose amplitude response is skew-symmetrical about ½ f_s, and whose group-delay remains constant over the frequency band of significant amplitude response.[6]

A class of filters that meets this criterion have a cosine roll-off of amplitude response with frequency. Fig. 2(a) shows the amplitude response of these filters for different rates of cut and Fig. 2(b) shows their respective impulse responses.[4] The impulse response has a single peak and is zero at spacings which are a multiple of the symbol interval from this peak. Hence, impulses can be sent at the symbol rate without interference between peaks.

The theoretical minimum signal bandwidth (½ f_s) is given by the 0% roll-off filter. However, it is impossible to make such a filter because it has an infinite rate of cut. Practical digital signalling systems normally use a pulse shape which occupies considerably more than the theoretical minimum bandwidth. The amplitude of the oscillatory tail of the impulse response decreases as the bandwidth of the signal is increased. As a result, the pulse shape becomes less sensitive to instrumental errors in the spectrum shaping filters. For lumped-element filters the less steep the cut-off that is required implies less variation in group-delay response and thus less equalisation is needed to avoid intersymbol interference. Taking the above arguments into account, the minimum bandwidth that can be used in a practical system is about 0·65 f_s for a baseband system and 1·3 f_s for a d.s.b. carrier system. This would correspond to a 30% cosine roll-off.

In the experimental modem, the overall spectrum shaping applied to the d.p.s.k. signal has a 100% cosine roll-off or "raised-cosine" form. Thus, the d.p.s.k. signal occupies twice the theoretical minimum bandwidth but the spectrum shaping filters are relatively simple to make.

2.5. Carrier recovery

For a random message, a 4-phase d.p.s.k. signal contains little energy at the carrier frequency and to regenerate a carrier from the signal, a non-linear process must be used.

There are various methods of carrier recovery, two of which will be briefly described here. A fuller discussion is given in an associated report.[7] All the methods rely on the fact that the d.p.s.k. signal spends much of its time in or near one of the nominal phase-states. Also, the amplitude of the signal is greatest at each of the four phase-states.

The first method for carrier recovery is known as the "phase-multiplication" method.[8] In essence, the 4-phase d.p.s.k. signal is passed through a network with a fourth order non-linearity so that the output contains, among many spectral components, a large component at four times the input carrier frequency. A voltage-controlled oscillator is phase-locked to this component and its output is frequency-divided by four to produce a reference signal at the original carrier frequency. The presence of the frequency divider means that the output carrier can have any one of four phases. For synchronous demodulation of a differentially-coded signal this four-phase ambiguity is of no consequence (see Section 2.3.).

The second or "remodulation" method[9] is illustrated in Fig. 3. The incoming d.p.s.k. signal is demodulated by the in-phase and quadrature components of a locally generated carrier. The demodulated signals are first limited and then used to remodulate quadrature components of the original d.p.s.k. signal. With the appropriate polarity of connections of the demodulated signals, the remodulation process tends to cancel the incoming modulation rather than double it. When the outputs of the remodulators are summed, a coherent component is produced at the original carrier frequency. The locally generated carrier is then phase-locked to this component.

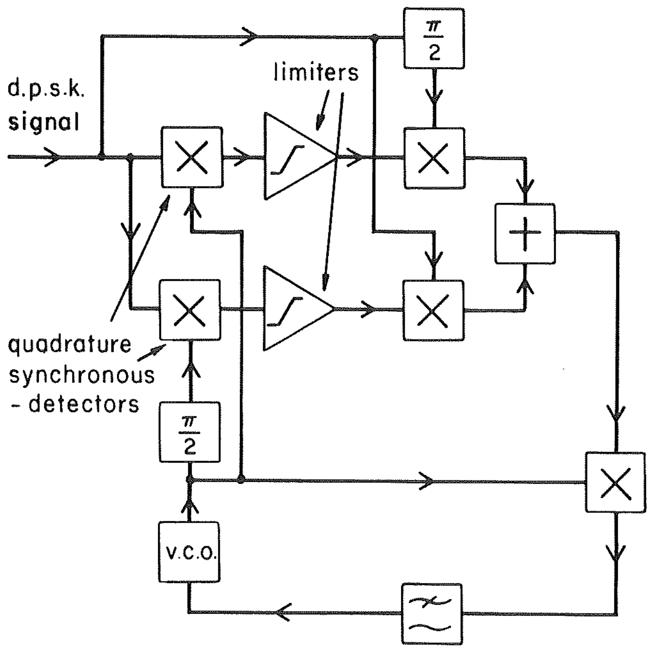


Fig. 3 - Carrier recovery by remodulation

In common with the first method, the recovered carrier has a four-phase ambiguity.

Various attempts[9,10] have been made to analyse these methods of carrier recovery under all possible signal conditions but none has been entirely successful. The non-linear nature of the signal-processing makes analysis almost impossible and computer simulation has often been used. It is relatively easy to prove that both methods will recover a phase-invariant carrier when the d.p.s.k. signal changes instantaneously between the four phase-states. However, for the case when the signal bandwidth is restricted so that transitions between phase states occupy a symbol interval the situation becomes much more complex. It is also relatively easy to prove that the recovered carrier will have four-phase ambiguity by applying arguments of symmetry.

Both methods of carrier recovery will work for all possible messages provided the total bandwidth allowed for the d.p.s.k. signal is greater than 1·5 times the symbol rate f_s. The worst-case d.p.s.k. signal for carrier recovery is one whose phase changes at the symbol rate in 90° steps. The frequency spectrum of such a signal which changes instantaneously between phase states is shown in Fig. 4; components at frequencies greater than $\pm f_s$ from the carrier frequency f_c have been ignored. There are only two components within $\pm f_s$ of the carrier, one at $f_c + (f_s)/(4)$, the other at $f_c - (3f_s)/(4)$. Both are required if a carrier is to be recovered. We must also remember that the reverse direction of stepping will give the mirror-image spectrum centred on f_c. If the total signal-bandwidth is less than $1{\cdot}5\, f_s$, the $f_c - (3f_s)/(4)$ component is removed and carrier recovery becomes impossible. This limiting condition can easily be avoided in practice. For example, most digital signals contain framing sequences within the data to aid decoding and these can be chosen to ensure that carrier recovery is possible for all messages.

In the experimental modem the demodulating carrier was recovered from the d.p.s.k. signal using the "phase-multiplication" method. At the start of this investigation,

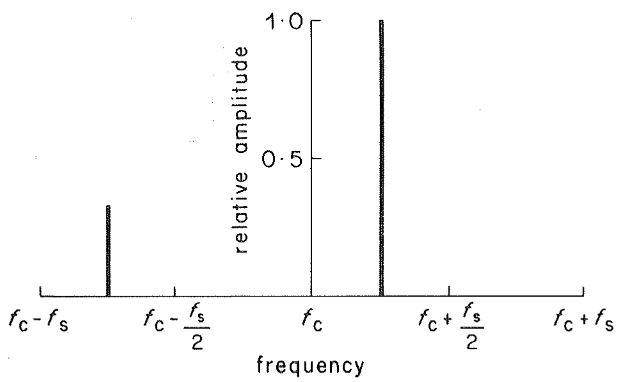


Fig. 4 - Frequency spectrum of 4-phase d.p.s.k. signal whose phase is 'switched' at the symbol-rate in 90° steps

f_c = carrier frequency
f_s = symbol-rate

the method was the best documented of the two. Also several researchers have reported instrumental difficulties with the "remodulation" method. More recent work within Research Department[7] and elsewhere, however, suggests that the "remodulation" method has several advantages some of which are discussed in Section 4.4.

2.6. Clock recovery

An essential part of the signal detection process is a clocking-signal, which samples the outputs of the two synchronous demodulators at the symbol rate. Each symbol of the signal must be sampled when its value has reached one of the discrete signal-states. Sampling during transitions between states will increase the probability of misinterpreting the message when noise is present.

A suitable clocking-signal can be recovered from the transitions of the demodulated signals themselves. As discussed in Section 2.3., the d.p.s.k. signal is synchronously demodulated by in-phase and quadrature components of a reference carrier. The outputs of the two demodulators are two-level pulse streams and the transitions between the two levels are separated in time by an integral number of symbol periods. If a circuit is arranged so that a "half-symbol-width" pulse is generated for every transition, then a coherent frequency component at the symbol-rate is produced. This train of "half-symbol-width" pulses can be filtered to give a phase-continuous clocking-signal.

2.7. The form of errors

When the 4-phase d.p.s.k. signal is demodulated at the receiver, unwanted noise accompanying the signal may cause some symbols to be misinterpreted. The transmitted message is coded as phase-changes between one symbol of the d.p.s.k. signal and the next; each phase-change carries two bits of information. A single symbol error will cause two phase-changes to be misinterpreted and consequently more than one bit will be in error. The form of errors is different from that found in two-level baseband systems in common use for digital audio distribution within the BBC. In the latter case, isolated single-bit errors are the most common, and multiple errors are quite rare. Error-protecting schemes that have been developed for present BBC digital sound transmission systems will have to be modified if they are to work in a 4-phase d.p.s.k. system.

The form of errors in a 4-phase d.p.s.k. system can easily be derived from first principles, by considering the sequence of events during signal detection.

In Fig. 5, the successive states of the d.p.s.k. signal for an arbitrary bit-stream are shown in vector form by the vectors a̲, b̲ and c̲. The phase difference between symbols 1 and 2 is 180° and between symbols 2 and 3 is −90°. The dotted lines represent the decision axes of the synchronous demodulation process. These decision axes enable the received phase of the d.p.s.k. signal to be placed in one of four quadrants. By examining the quadrants occupied by successive symbols, the phase difference between symbols can be deduced and the message

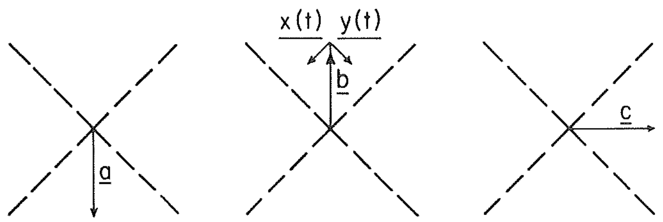


Fig. 5 - Vector diagram showing successive symbols of the d.p.s.k. signal

decoded.

Let us assume that symbols 1 and 3 are correctly detected, but that symbol 2 is misinterpreted because of noise on the d.p.s.k. signal. For convenience, the added noise has been resolved into two statistically independent quadrature components x(t) and y(t) of equal r.m.s. value (see Fig. 5). Symbol 2 will be placed in one of the adjacent quadrants if either of the quadrature noise components crosses one of the decision axes. If the probability of one of the noise components crossing a decision axis is p then provided this is small, the probability of symbol 2 being placed in either one of the adjacent quadrants is $p + p = 2p$. For symbol 2 to be placed in the opposite quadrant, both of the quadrature noise components must cross the decision axes simultaneously. The probability of this event is therefore $p \times p = p^2$.

The placing of a symbol of the d.p.s.k. signal in one of the adjacent quadrants is thus by far the most likely form of error due to random noise. Placing symbol 2 in one of the adjacent quadrants will cause the phase difference between symbols 1 and 2 to be detected with $\pm 90^{\circ}$ error. A similar error will occur for the phase difference between symbols 2 and 3.

The number of bits that will be decoded in error, for a single symbol error, depends on the way in which the d.p.s.k. signal is coded. As described in Section 2.3., the transmitted data signal is split into pairs of digits and each pair changes the phase of the signal by a preset amount. Figs. 6(a) and 6(b) show two possible ways of assigning the four possible pairs of digits to the transmitted phase-changes. Using the format of Fig. 6(a), if a single phase-change is detected by the receiver with a $\pm 90^{\circ}$ error, two bits can be in error. For the cyclic format (Gray code) of Fig. 6(b) only one bit can be in error.

If the d.p.s.k. signal is coded using the cyclic format, the incidence of bit errors can be summarised as follows. Whenever a symbol is placed erroneously in one of the adjacent quadrants, two successive phase-changes will be detected with a $\pm 90^{\circ}$ error. As a result, two bits within a group of four bits will be decoded in error. On rare occasions when a symbol is placed in the opposite quadrant, all four bits within the group will be in error (see Fig. 6(b)).

3. Experimental circuits

3.1. General description

A block schematic diagram of the modem is shown in Fig. 7. It operates at an intermediate frequency of 10·7 MHz; the standard i.f. was chosen for convenience.

The following sections describe each of the units of the modem in some detail. The basic system parameters and the function of each of the units have been discussed in Section 2.

3.2. DPSK modulator

The modulator has two distinct parts. Analogue circuitry operating at 10·7 MHz produces the 4-phase signal by rapid switching of the phase of the output from

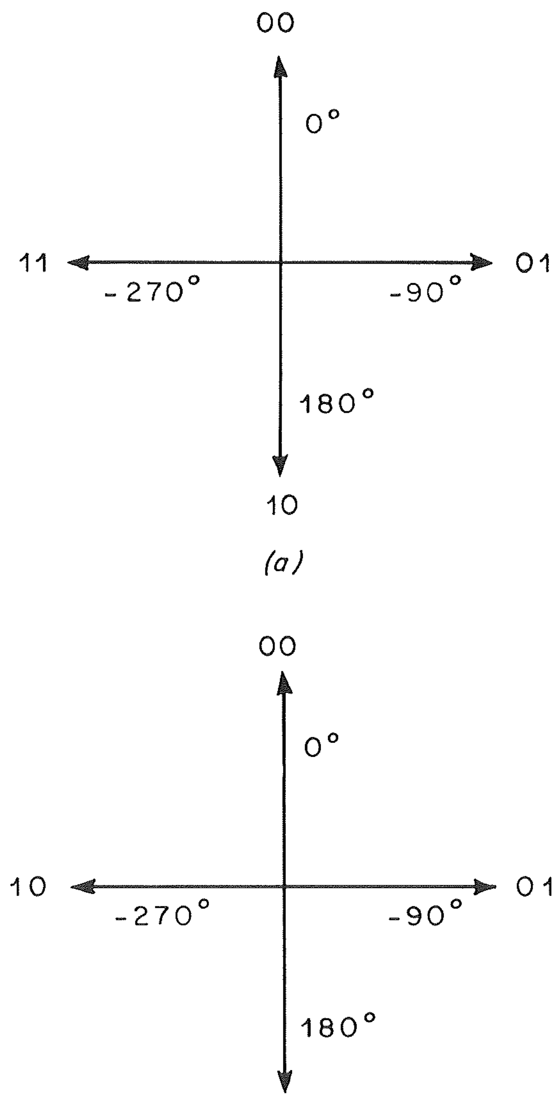


Fig. 6 - Two possible ways of assigning the four possible pairs of digits to the transmitted phase-changes

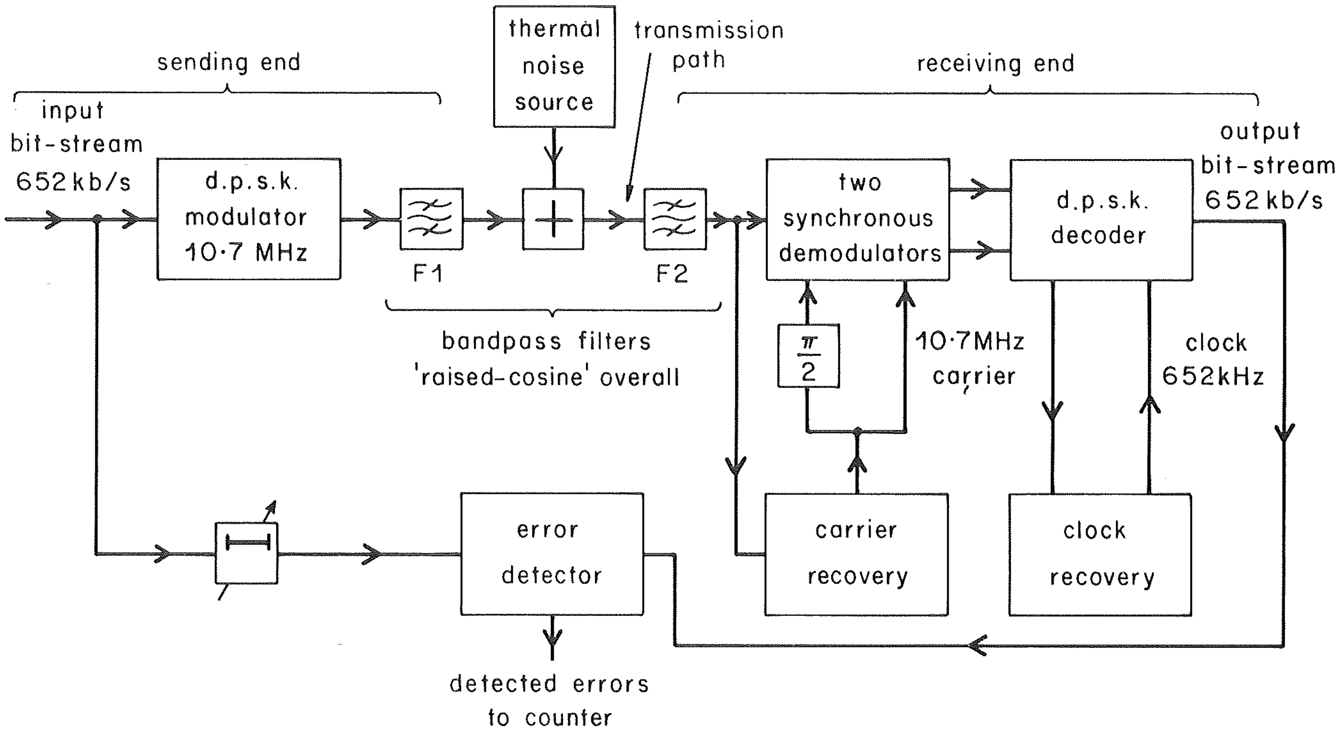
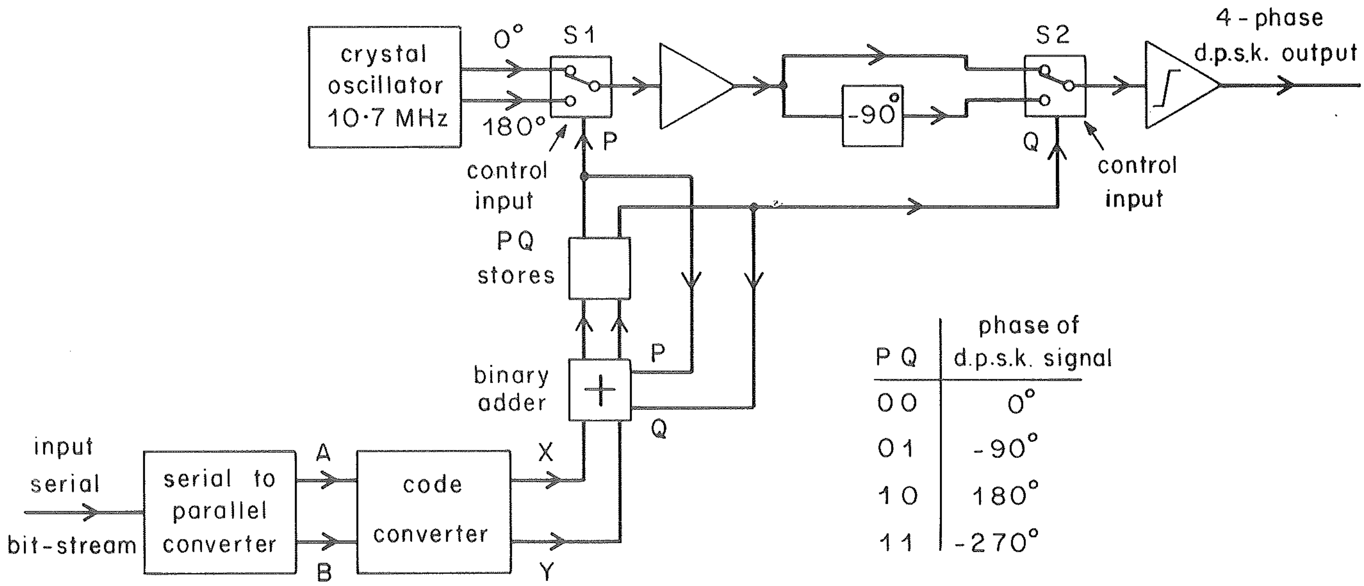


Fig. 7 - Experimental modem and test arrangements

a crystal-oscillator. The output signal is identical to that produced by the amplitude-modulation of in-phase and quadrature carriers with rectangular pulses. Digital circuitry converts the incoming serial bit-stream into a form suitable for driving the switches that set the phase of the d.p.s.k. signal.

With reference to Fig. 8(a), the 10·7 MHz crystal-oscillator has two outputs with a phase-difference of 180° between them and electronic changeover switch S1 selects one of the two outputs. When the control input to switch S1 is a logical '0', the 0°-phase of carrier is selected and the 180°-phase when a logical '1'. The output from S1 is passed through a wideband buffer amplifier before passing to the next stage. Here the changeover switch S2 selects the 0°-phase of carrier when the control input is a logical '0' and the -90°-phase when a logical '1'. The -90° phase-shifter is an all-pass lumped-element network.

The output of switch S2 is limited before leaving the modulator to reduce any amplitude differences between



PQ	phase of d.p.s.k. signal
00	0°
01	-90°
10	180°
11	-270°

(a) *Fig. 8 - DPSK modulator* (b)

the four phases of the d.p.s.k. signal.

Switch S1 is identical to switch S2; both are made with Complementary-Metal-Oxide-Silicon (CMOS) analogue multiplexers. The switches will pass signals up to 40 MHz with little distortion, and have a changeover time of approximately 100 ns.

Fig. 8(b) shows the phase of the d.p.s.k. signal as a function of the state of the switch-control inputs P and Q. PQ can be considered as a binary number with P as the most significant digit. If a two-bit binary number is added to PQ, then the phase of the modulator output signal always changes by the appropriate amount, independent of the value PQ. For example, if 01 is added to PQ, the phase changes by -90°. The message sent by the modulator is coded as a difference in phase between adjacent symbols of the d.p.s.k. signal and, hence, any pair of digits can be sent simply by adding them to the binary number PQ.

The input serial bit-stream to the modulator is split into pairs of digits shown as A and B in Fig. 8(a) by a serial-to-parallel converter. The 4-phase d.p.s.k. modulator transmits two bits of information per symbol. Each pair of digits A and B is passed through a code converter whose outputs X and Y determine the phase-change to be sent by the modulator. The code conversion can be varied and all transformations between two-bit binary numbers are possible. The code converter is normally set so that the d.p.s.k. signal is coded using the cyclic format shown in Fig. 1.

The digits X and Y, which represent the phase-change to be transmitted, are added to PQ in binary form to produce a new value for PQ and hence a new phase for the output signal.

3.3. Spectrum shaping filters

As discussed in Section 2.4., the modem has been designed so that the demodulated d.p.s.k. signals have a "raised-cosine" spectrum. The filters that give this spectrum are shown as F1 and F2 in Fig. 7. Both the filters are bandpass and operate at the intermediate frequency of 10·7 MHz. In any practical modulation system, there will be other filters in the radio-frequency circuits of the transmitter and receiver. These filters should have little effect on the spectrum shaping and they will be ignored.

The spectrum shaping can be divided in many ways between the transmitter and receiver. Where average signal power is the limiting factor, minimum error-rate in the presence of noise is obtained when the spectrum shaping is divided equally between the transmitter and receiver. This assumes that the source signal is produced by modulating quadrature carriers with impulses. The optimum is not very critical and quite large departures from it cause little degradation. In the experimental modem, part of the spectrum shaping is done by the modulator itself, and the signal from the modulator is the same as that produced by modulating quadrature carriers with rectangular pulses rather than impulses. For convenience, the remainder of the spectrum shaping is concentrated at the receiver. The role of the transmitter filter is merely to reject out of band components from the modulator.

The source filter F_1, has a rectangular amplitude response which restricts the bandwidth of the d.p.s.k. signal to ± 326 kHz. The signal at the output of the modulator can be regarded as changing instantaneously between phase states, and it has a spectrum of the form (sin x)/(x) with zero energy at frequencies spaced from the carrier by multiples of the symbol-rate (326 kHz). The filter F_1 is a 0.1 dB ripple 4-section Tchebycheff filter with a ripple bandwidth of ± 300 kHz. The filter was transformed from its low-pass equivalent using the Cohn transformation.[11] It was realised that, using this transformation, the resulting response would be geometrically symmetrical rather than arithmetically symmetrical about the carrier frequency. However the percentage bandwidth of the filter (~6%) is so small that this effect can be neglected.

The receiver filter F_2 converts the truncated (sin x)/(x) spectrum at the output of the transmitter filter into a "raised-cosine" spectrum. The ideal amplitude response and the measured response of the experimental filter are shown in Fig. 9(a). The response of the experimental filter closely follows the ideal over the central part of the curve. The d.p.s.k. signal has little energy outside this central region, and so the error in the response of the experimental filter causes little intersymbol interference.

The design of the receiver filter was entirely empirical. The desired response can be approximated by constructing a standard filter type such as Butterworth or Tchebycheff with lossy components. By examining graphs of the responses of such filters whose inductors have a finite Q-factor, a suitable filter was designed. The experimental filter is based on the design of a 3-section Tchebycheff filter with a ripple bandwidth of ± 163 kHz, and the inductors have a Q-factor of about 80. In common with filter F_1, the design was transformed from its low-pass equivalent using the Cohn transformation.[11] During its construction various trial inductors were made with different ferrite cores and wire gauges until the correct Q-factor was found.

The spectrum shaping filters F_1 and F_2 were designed to give the correct overall amplitude response without taking account of their group-delay responses. For no intersymbol interference, the group-delay of the filters should be constant within the signal bandwidth.

For simplicity, the group-delay responses of both the filters were equalised by a single unit. The group-delay response of the filters in tandem is shown in Fig. 9(b). The response has a variation within the passband of 0·5 μs, and it is asymmetric about the carrier frequency. To equalise the group-delay response a 3-section active equaliser was built. Each of the active sections[12] has a single peak of group-delay whose centre frequency and Q-factor can be varied independently.

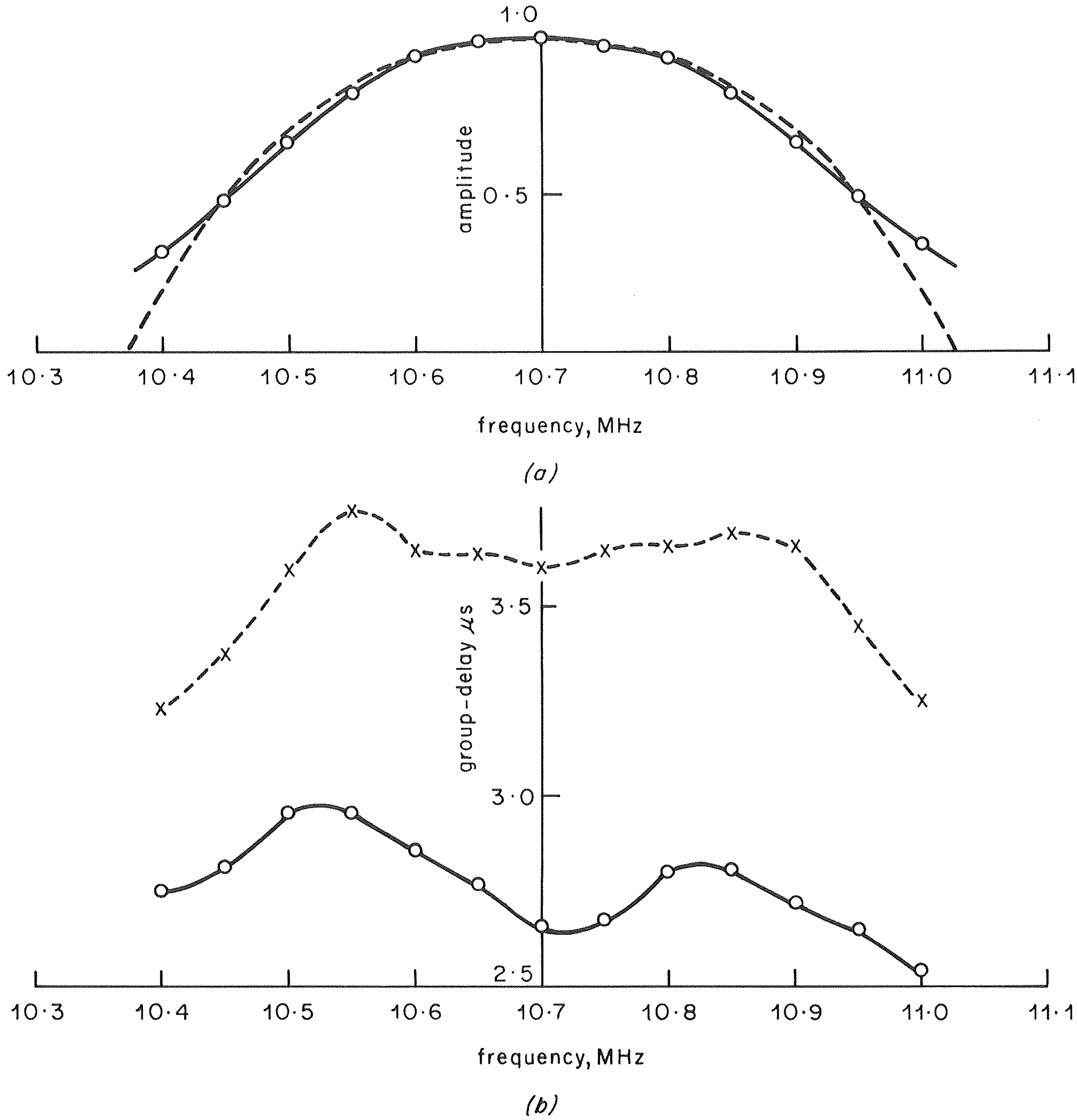


Fig. 9 - Amplitude and group-delay responses of spectrum shaping filters

(a) amplitude response of receiver filter F_2

------ ideal

o———o measured

(b) group-delay response of receiver filter F_2 and transmitter filter F_1 in tandem

o———o before equalisation

x------x after equalisation

In a short experiment, the equaliser sections were brought into operation in turn to equalise the group-delay response of the filters. Before adding the next section of the equaliser, the level of intersymbol interference on the demodulated data signals was observed. During this experiment it was found that, for minimum intersymbol interference, symmetry of the group-delay response about the centre-frequency was the most important factor. Symmetrical departures from a flat group-delay response of up to 0·5 μs produced very little intersymbol interference. The effect of symmetrical errors in group-delay response can be found by considering their effect on a baseband data signal. Sunde[13] has shown that "raised-cosine" pulses are relatively tolerant of quite large variations in group-delay response. On the other hand, small assymetric errors in group-delay response cause large amounts of crosstalk between the quadrature axes of a quadrature modulated signal.[14] In a 4-phase p.s.k. or d.p.s.k. system this crosstalk results in interference between the data signals carried on each quadrature axis.

In the final group-delay equaliser, only one of the three original equaliser sections is used, and the equalised group-delay response is shown in Fig. 9(b). The overall variation of the group-delay response within the passband is about the same as before equalisation, but the response is much more symmetrical.

3.4. Carrier recovery circuit

As discussed in Section 2.5., the method of carrier recovery is the "phase-multiplication" method.

Referring to Fig. 10, the d.p.s.k. signal is passed through a network with a fourth order non-linearity formed from two integrated circuit square-law devices in tandem. These devices do not have a perfect square-law

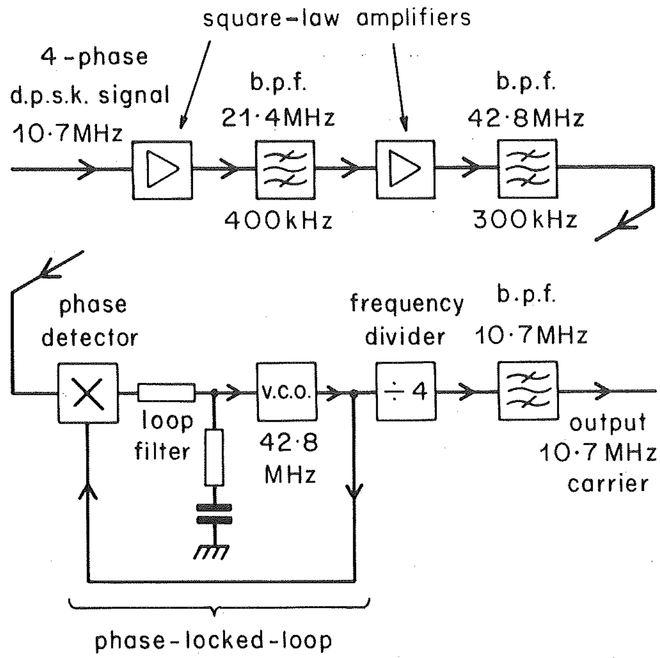


Fig. 10 - Carrier recovery circuit

so that their outputs contain unwanted harmonics of the input signal which are removed by the 21·4 MHz and 42·8 MHz bandpass filters.

The spectrum of the output from the 42·8 MHz b.p.f. for a randomly modulated d.p.s.k. signal is shown in Fig. 11. Besides the component at four times the input carrier frequency there are strong components spaced at the symbol-rate frequency (326 kHz).

The four-times-carrier-frequency component is ex-

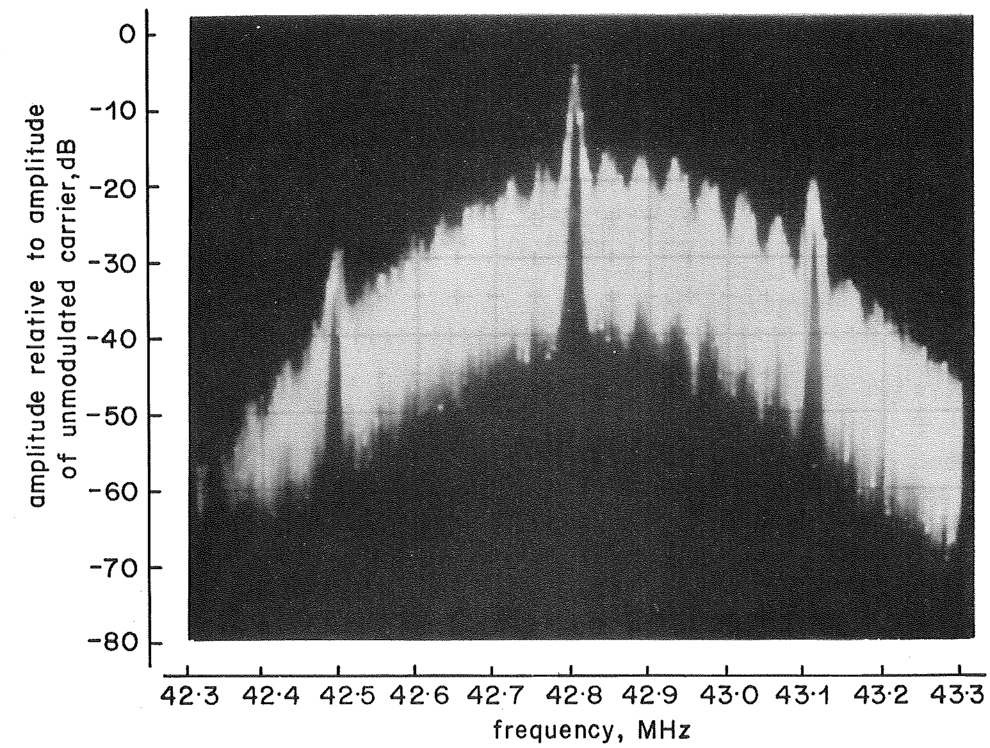


Fig. 11 - Spectrum of output from four-times-frequency multiplier, analyser bandwidth 10 kHz

tracted using a phase-locked-loop[15] (p.l.l.). The p.l.l. acts as a narrow-bandwidth bandpass filter whose centre-frequency tracks variations in the frequency of the input signal. The tracking ability of the p.l.l. is set by the d.c. gain of the loop. The loop-filter controls the noise-bandwidth and the dynamic performance of the loop. The type of filter shown in Fig. 10 gives the loop a second-order response.

The design of the p.l.l. is complicated by the fact that the amplitude of the four-times-carrier-frequency component can vary over a 6 to 1 range. The amplitude of this component controls many of the parameters of the p.l.l. For example, the d.c.-gain of the loop is proportional to the amplitude. Hence the noise-bandwidth, pull-in range etc. vary, depending on the form of the d.p.s.k. signal. The amplitude is maximum when the phase of d.p.s.k. signal remains constant and minimum when the phase changes at the symbol-rate in 90° steps. For a randomly modulated signal the amplitude is about half the maximum value. Care has been taken to ensure that the p.l.l. performs adequately under all signal conditions.

The main design parameters of the p.l.l. are listed below. All the parameters are quoted for the p.l.l. itself, not the complete carrier-recovery circuit.

	Best-case value	Worst-case value
Pull-in range	300 kHz	50 kHz
Phase-error for 50 kHz change in input frequency	$^{2}/_{3}{}^{\circ}$	4°
Single-sided noise-bandwidth	2·3 kHz	7·4 kHz

The voltage-controlled-oscillator output is frequency-divided by four, using two Schottky TTL D-Type flip-flops. The recovered carrier at 10·7 MHz is filtered to remove unwanted harmonics and the output signal is then used to synchronously demodulate the d.p.s.k. signal.

3.5. Synchronous demodulators and decoding logic

The received d.p.s.k. signal is synchronously demodulated by in-phase and quadrature components of the recovered reference carrier as shown in Fig. 12(a). The outputs P' and Q' of the demodulators are "raised-cosine" pulse streams about earth potential. Fig. 12(b) is a vector diagram of the synchronous demodulation process and Fig. 12(c) shows the polarity of the signals P' and Q' as a function of the 'quadrants' of the d.p.s.k. signal. The synchronous detectors are integrated-circuit types with a single-sided modulation-bandwidth of 2 MHz which greatly exceeds the single-sided bandwidth of the d.p.s.k. signal (326 kHz).

The "raised-cosine" pulse streams P' and Q' are sampled at the symbol-rate by level-comparators and D-Type flip-flops. The outputs of the flip-flops indicate the polarity of the signals P' and Q' at the sampling instant.

The remaining signal processing is done in the digital domain using CMOS integrated circuits. The processing takes the form of several transcoding operations and may appear unnecessarily complicated but the instrumentation is relatively simple. Much of the processing is the inverse of that in the modulator. Many of the signals in the demodulator have an exact equivalent in the modulator

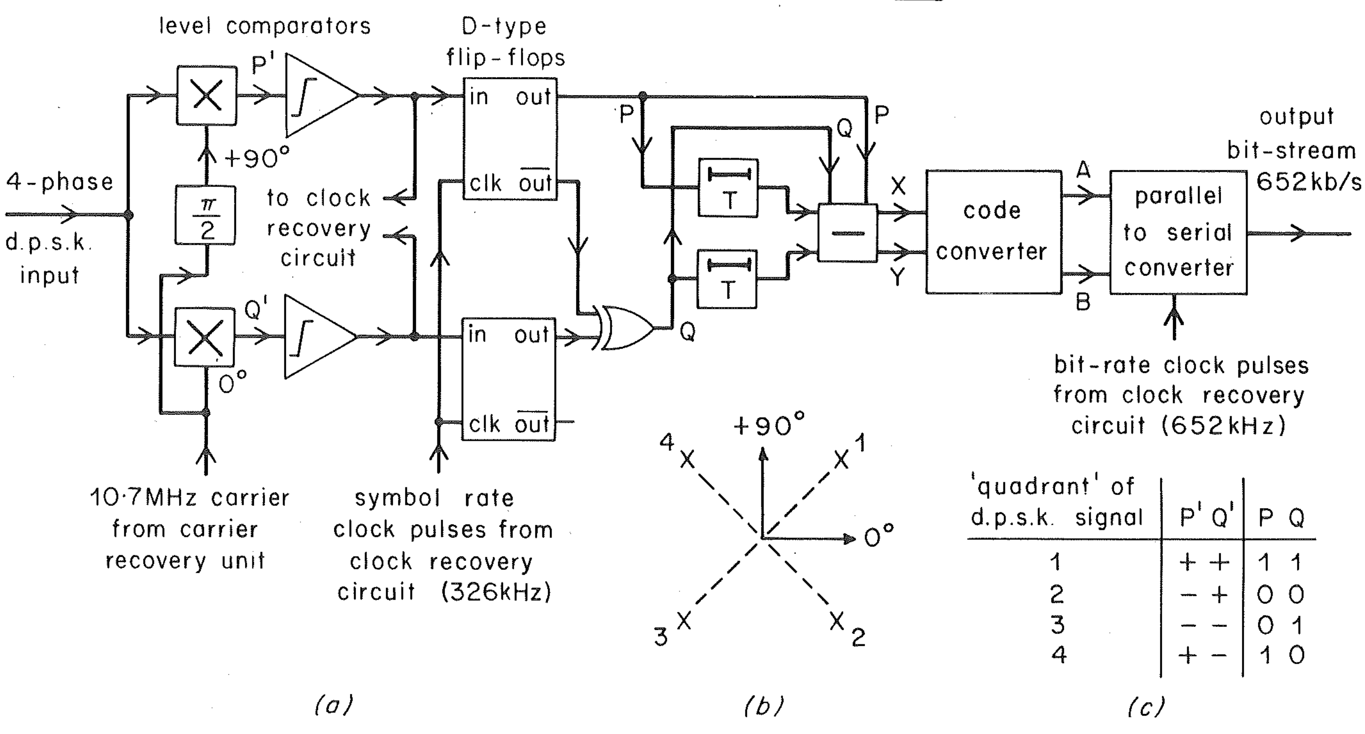


Fig. 12 - Synchronous demodulation and decoding, T = one symbol period delay

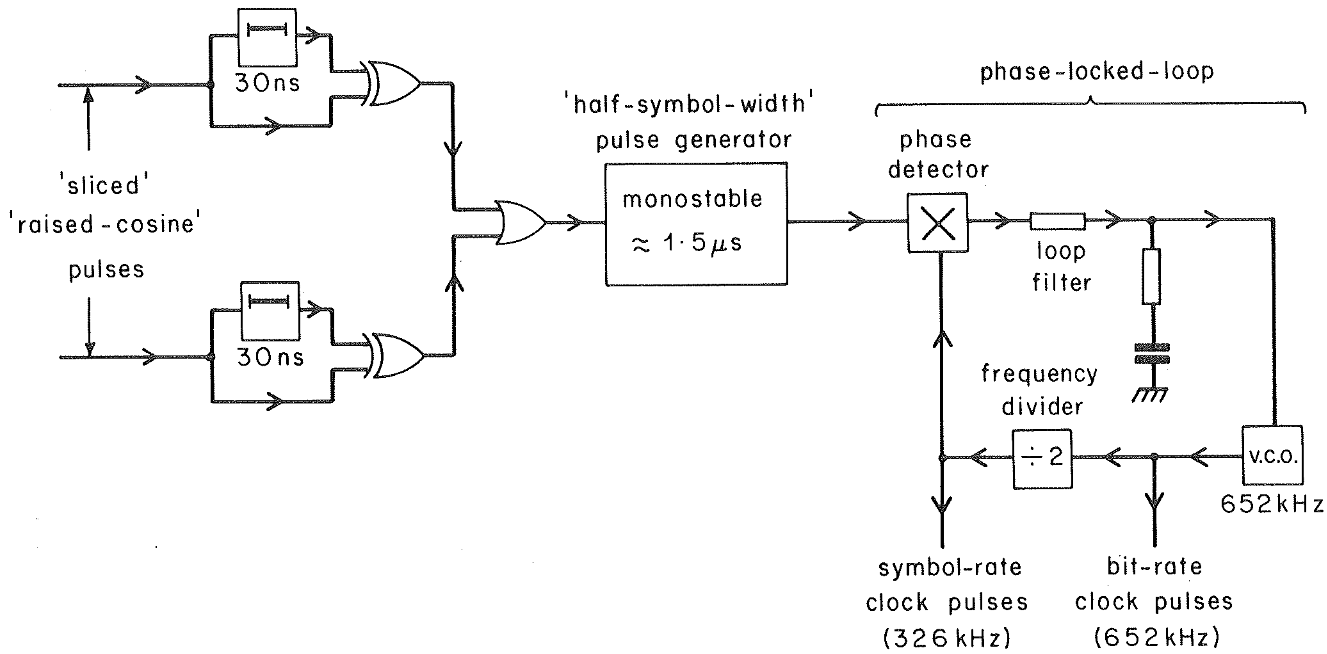


Fig. 13 - Clock recovery circuit

and are denoted by the same letters.

The sampled values of P′ and Q′ are converted into P and Q as shown in Fig. 12(a) and the relationship between PQ and the "quadrants" of the d.p.s.k. signal is shown in Fig. 12(c). PQ can be regarded as a two-bit binary number with P as the most significant digit. As the phase of the d.p.s.k. signal changes in -90° steps (phase rotating clockwise in Fig. 12(b)) the value of PQ increases in binary steps. Hence the phase-change between adjacent symbols of the d.p.s.k. signal can be found simply by subtracting the previous value of PQ from the present value of PQ. The result of this subtraction is a two-bit binary number XY.

The digits XY are passed through a code-converter whose outputs A and B are the original input data to the d.p.s.k. modulator. The code converter in the demodulator is the inverse of that in the modulator and the two must be pre-set to be complementary. The signals A and B are re-formed into a serial bit-stream which is identical to the bit-stream at the input to the modulator.

3.6. Clock recovery circuit

As discussed in Section 2.6., a clocking-signal is regenerated from the transitions of the demodulated data signals. The method is described in detail in Reference 16 only a brief description is given here. A block schematic diagram of the circuit is shown in Fig. 13.

The circuit receives the demodulated data signals sliced midway between the two extremes of level and the resulting transitions are detected by two exclusive-OR gates. The outputs from these gates are 30 ns-wide pulses which trigger the monostable to produce "half-symbol-width" (~1·5 μs) pulses.

The train of half-symbol-width pulses has, within its wide spectrum, a coherent component at the symbol-rate. The amplitude of the symbol-rate component is proportional to the probability of a transition in the demodulated data signals. This probability is in turn related to the audio signal being carried by the system. When the audio channels are active there will be a greater probability of a transition than when the channels are quiescent and the p.c.m. bit-stream is mostly zeros. It was anticipated that the probability of a transition would vary over an approximately 4 to 1 range.

The symbol-rate component in the stream of half-symbol-width pulses is extracted by a p.l.l. A voltage-controlled-oscillator with a free-running frequency close to the bit-rate (652 kHz) is phase-locked to the pulse stream. In common with the p.l.l. in the carrier recovery circuit, the performance of the clock recovery p.l.l. depends on the form of the transmitted message. Care has been taken to ensure that the p.l.l. gives the desired performance under all anticipated signal conditions.

The main design parameters of the p.l.l. are listed below. The parameters are quoted assuming that the voltage-controlled-oscillator and frequency-divider are lumped together to give an effective oscillator frequency of 326 kHz.

	Best-case value	Worst-case value
Pull-in range	60 kHz	15 kHz

	Best-case value	Worst-case value
Single-sided noise-bandwidth	3·3 kHz	9·9 kHz
Phase-error for 10 kHz change in input frequency	2·5°	10°

The recovered symbol-rate clocking-signal is passed to the demodulator where it samples the raised-cosine pulse streams P′ and Q′ (Fig. 12). The recovered bit-rate clocking-signal is passed to the decoder and is used to re-form the demodulated data signal into a serial bit-stream.

4. Performance of experimental modem

4.1. Error rate versus carrier-to-noise ratio

A block schematic diagram of the experimental arrangements for measuring the noise performance of the modem is shown in Fig. 7. The digital signal used in the tests was a long pseudo-random sequence which repeated itself every 2^{15} - 1 bits. The demodulated sequence was compared with a suitably delayed version of the source sequence and the detected errors were counted over a period of up to 100 seconds.

Wideband Gaussian noise was added to the d.p.s.k. signal between the two spectrum shaping filters. The noise source was a Zener diode whose output had to be amplified by 50 dB to give equal carrier and noise powers at the input to the demodulator. Care was taken to ensure that noise peaks did not overload the amplifiers and so distort the Gaussian properties of the noise.

The carrier power and noise power at the input to the demodulator were measured using a bolometer type power meter. The carrier-to-noise ratio could be varied in 0·1 dB steps over the range 0 to 25 dB, with an accuracy of ± 0·2 dB.

The results of the noise performance tests on the modem are summarised in Fig. 14. The modem performs within approximately 0·5 dB of the theoretical limit.[4] At error rates below 1 in 10^7 errors occur too infrequently for accurate measurement.

For a 10-bit p.c.m. sound signal using a near-instantaneous compandor, without error protection except for the scale-factor word, the threshold of error perceptibility is at about 1 in 10^7 error rate.[17] This corresponds to a carrier-to-noise ratio of 15 dB for the d.p.s.k. signal. When a simple error-protection scheme is used, the threshold of perceptibility should be raised to an error-rate of about 1 in 10^5,[17] corresponding to a carrier-to-noise ratio of 13·4 dB. The latter figure will have to be verified by subjective assessment of the audio signal, since the errors in a d.p.s.k. system are somewhat different in form from the random errors which were used in the experiments cited.

In the following sections, the performance of individual units within the modem is discussed and their effect on the performance of the modem is assessed.

4.2. DPSK modulator

Ideally, the phase difference between any of the four possible phases of the modulator output should be either ± 90° or 180°. The accuracy of the four phases can then be specified as the departure of the measured phase differences between phase-states and the ideal. Since the data transmitted by the modulator is coded as carrier phase-changes, no particular phase state can be used as a reference phase.

In a short experiment, the phase difference between any two states was measured with a vector voltmeter when the modulator was switched manually between states; one output from the carrier oscillator was used as the reference for the vector voltmeter (see Fig. 8). It was found that the phase-shifters within the modulator could easily be set so that the phase difference between states was within 1° of the ideal. Such a small error has negligible effect on the performance of the modem.

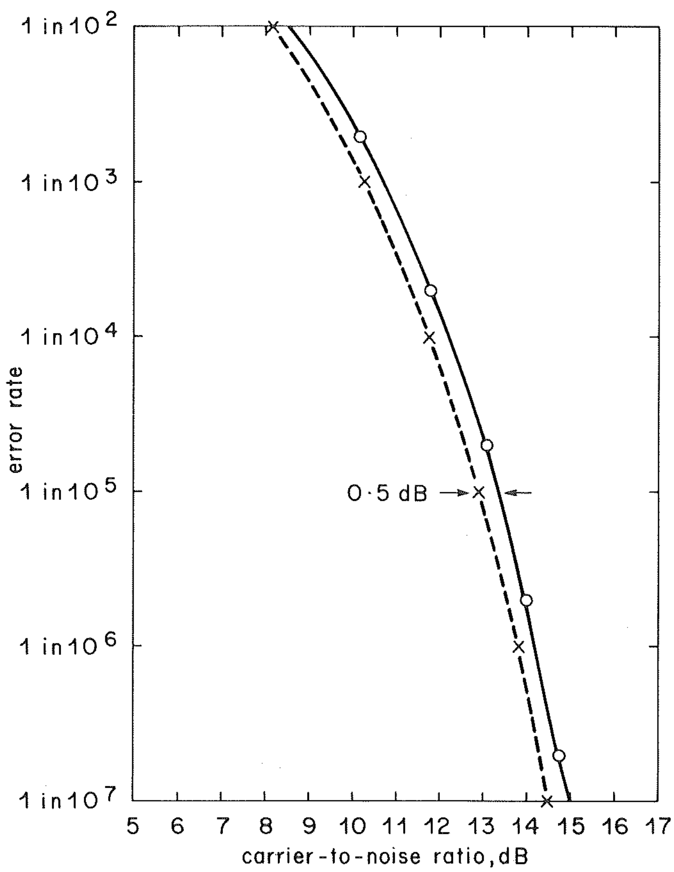


Fig. 14 - Error rate versus carrier-to-noise ratio

o——o = measured
x- - - -x = theoretical

No noticeable drift in the four phases of the signal was found when the equipment was working in the laboratory. The phase of the signal is determined largely by passive phase-shifters whose temperature stability is high enough to be ignored.

4.3. Spectrum shaping filters

The characteristics of the spectrum shaping filters determine the level of intersymbol interference on the demodulated data signals.

A photograph of the demodulated data signals displayed on an oscilloscope for a pseudo-random message is shown in Fig. 15. The horizontal time base for the oscilloscope was triggered by the recovered symbol-rate clocking-signal. A convenient method for specifying the level of intersymbol interference on these signals uses the concept of "eye-height", a term that refers to the eye-like shape of the central portion of the waveforms shown in the photograph. The eye-height is defined as the ratio of the minimum separation between the two signal-levels at the centre of the eye to their separation when the signals are all '1's or all '0's. The eye-height was measured as approximately 97% which reduces the noise margin of the modem by 0·3 dB.[4]

As discussed in Section 3.3., the eye-height was very sensitive to asymmetry in the group-delay response of the spectrum shaping filters. Prior to group-delay equalisation the filters gave an eye-height of approximately 90% which would have degraded the noise margin of the modem by 0·8 dB.[4]

4.4. Carrier recovery circuit

Tests confirmed that the performance of the phase-locked-loop within the carrier recovery circuit closely followed theoretical predictions. The loop pulled into lock under all signal conditions and tracked variations in carrier frequency with the designed phase accuracy. The loop was found to be relatively insensitive to temperature variation.

It was noticed, however, that the phase of the recovered 10·7 MHz carrier drifted over a 3 to 4^{o} range during the warm-up period of the circuit. When the circuit had been working for an hour, the phase of the recovered carrier was stable within 1^{o}. Before each test on the modem, the carrier phase was adjusted to be correct. The most temperature-sensitive part of the circuit was found to be the four-times frequency-multiplier and its associated bandpass filters. In a practical system the circuits temperature sensitivity would have to be reduced. This might involve placing it in a temperature-controlled oven or using temperature-compensating elements.

A better solution to this problem would be to use the "remodulation" method of carrier recovery shown in Fig. 3; in this method the synchronous demodulators for the d.p.s.k. signal are within the system loop. Therefore the circuit will correct any phase error in the demodulating carrier caused by changes in component values with temperature. Initial doubts about instrumenting this method have been dispelled since it has been shown[18] that there is a simpler baseband equivalent which does not require actual remodulation of the incoming signal. A carrier recovery circuit using the "baseband remodulation" method has now been developed.[7]

The phase jitter on the recovered carrier was measured for different input carrier-to-noise ratios and the results are shown in Fig. 16. When the error rate of the

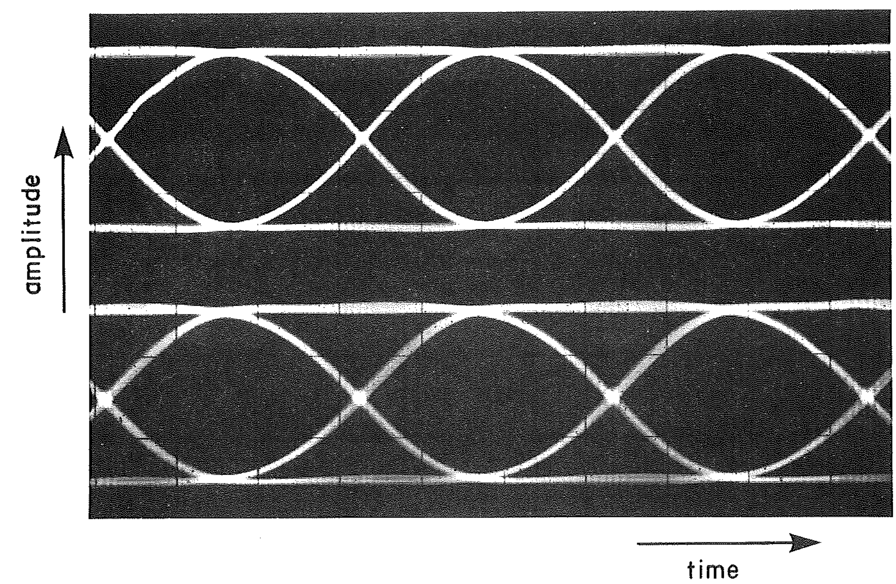


Fig. 15 - Outputs of synchronous demodulators

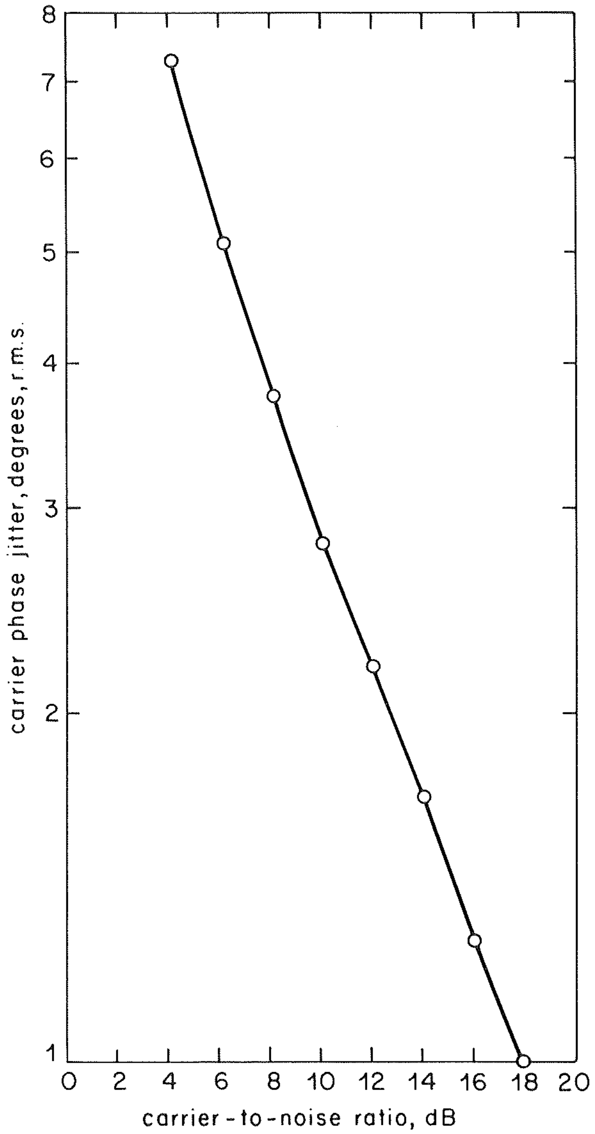


Fig. 16 - Phase jitter on recovered carrier versus carrier-to-noise ratio

demodulated data signals is 1 in 10^5 corresponding to a carrier-to-noise ratio of 13·4 dB the phase jitter is 1·8° r.m.s. In the Appendix it is shown that this level of jitter reduces the noise margin of the modem by 0·1 dB.

When the carrier-to-noise ratio falls below 4 dB, the phase-locked-loop within the circuit begins to "skip cycles".[15] The p.l.l. falls momentarily out of lock and then re-locks in a different phase from before. At these low carrier-to-noise ratios, noise peaks will often exceed the amplitude of the four-times-carrier-frequency component at the input to the p.l.l. If a noise peak inverts the phase of this component the polarity of the feedback within the p.l.l. is also inverted. If the condition persists, the p.l.l. is driven out of lock. When the noise peak subsides it will regain synchronism.

4.5. Demodulator sampling circuits

The circuitry that samples the raised-cosine pulse streams P′ and Q′ (see Fig. 12) does not have a very well-defined decision-threshold. Ideally, the sampler should be able to determine the polarity of the signals P′ and Q′ without error and hence have a single decision-threshold. In the practical circuit, the level-comparators need a small amount of overdrive before they can change state and hence the sampler exhibits an effect similar to hysteresis. As a result, the decision threshold is blurred and is described as having a finite width.

The effective threshold-width of the experimental sampling circuits was measured as 3% of the peak-to-peak amplitude of the raised-cosine pulse streams P′ and Q′. This threshold-width may be considered small, as it has little effect on the performance of the modem.

4.6. Clock recovery circuit

The performance of the clock recovery circuit closely follows theoretical predictions. The phase-locked-loop pulls into lock under all anticipated signal conditions. Changes in temperature have little effect on the phase of the recovered clocking-signal.

The phase jitter on the symbol-rate clock pulses was measured for different d.p.s.k. signal carrier-to-noise ratios and the results are shown in Fig. 17. When the bit-error rate is 1 in 10^5 corresponding to a carrier-to-noise ratio of 13·4 dB, the r.m.s. phase jitter is 6° which produces negligible impairment to the noise margin of the modem.

The limit on the amplitude of clock pulse jitter is set by the audibility of the jitter on the demodulated

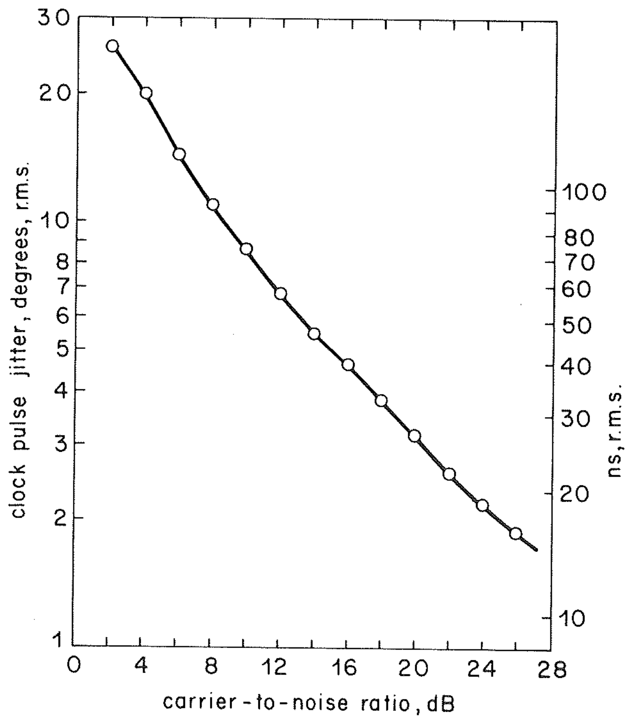


Fig. 17 - Phase jitter on recovered symbol-rate clock pulses versus carrier-to-noise ratio

audio signals. It has been shown that 50 ns r.m.s. of random timing jitter[19] is just perceptible to the most critical observers. This amplitude of jitter is exceeded when the carrier-to-noise ratio falls below 14 dB. This is also the point when digit errors become perceptible. To ensure that the jitter does not impair the audio signals, its level can be reduced by decreasing the noise-bandwidth of the clock recovery circuit. In the experimental circuit, this can be done relatively simply by using a crystal within the voltage-controlled-oscillator of the p.l.l. When the bit-rate for the modem has been accurately defined a suitable crystal will be inserted.

The p.l.l. begins to "skip-cycles" when the carrier-to-noise ratio falls below 2 dB. It is interesting to note that the clock recovery circuit can synchronise whilst the carrier recovery circuit is unlocked.

5. Conclusions

An experimental 4-phase differential-phase-shift-keying system has been constructed to explore its suitability for the transmission of two high-quality digital sound signals within a radio frequency bandwidth of approximately 650 kHz. The system offers a good compromise between the requirements of bandwidth, carrier power and instrumental simplicity.

Measurements have shown that the noise performance of the experimental equipment is within about 0·5 dB of the theoretical limit. The slight loss of performance is largely due to instrumental imperfections in two areas of the equipment. Firstly, the form of spectrum shaping applied to the signal is not perfect. The design of the spectrum shaping filters is quite critical if optimum performance is to be achieved. Secondly, the carrier which synchronously demodulates the received signal has a small amount of phase jitter. Although the experimental method of carrier recovery performs satisfactorily in this respect other methods could be used to advantage. The performance of another method is presently being studied.

In practical applications of the system, the performance will be degraded by other transmission path impairments such as multipath propagation, adjacent and co-channel interference. For terrestrial applications such as outside-broadcast links (as opposed to its use in satellite systems) multipath propagation is likely to be the most significant. The susceptibility of the experimental system to multipath propagation is currently under investigation. Only when this work is complete, and the effect of other transmission path impairments has been considered, will it be possible to decide what system refinements may be desirable and, if the results are encouraging, make recommendations in terms of a fully specified system.

6. References

1. SHORTER, D.E.L. *and* CHEW, J.R. 1972. Application of pulse-code modulation to sound-signal distribution in a broadcasting network. Proc. Instn. Elect. Engrs., 1972, 119, 10, pp 1442-1448.

2. OSBORNE, D.W. *and* CROLL, M.G. 1973. Digital sound signals: bit-rate reduction using an experimental digital compandor. BBC Research Department Report No. 1973/41.

3. CROLL, M.G., OSBORNE, D.W. *and* REID, D.F. 1973. Digital sound signals: multiplexing six high-quality sound channels for transmission at a bit-rate of 2·048 Mb/s. BBC Research Department Report No. 1973/42.

4. BENNETT, W.R. *and* DAVEY, J.R. 1965. Data Transmission, McGraw-Hill, 1965.

5. HARVEY, R.V. 1975. An experimental 8-phase modulation system for distribution of digital television signals. BBC Research Department Report No. 1975/10.

6. NYQUIST, H. 1928. Certain topics in telegraph transmission theory, Trans. AIEE, 47, April 1928.

7. MAHADEVA, W.A. A 4-phase differential-phase-shift-keying stereo sound system: carrier recovery methods. BBC Research Department Report in preparation.

8. ADAMS, G.W. 1969. A modem for the Datel 2400 service – Datel modem No. 7A, POEEJ, 62, 3, p 156, October 1969.

9. NOZAKA, K., MURATANI, T., OGI, M. *and* SHOJI, T. 1969. Carrier synchronisation techniques of p.s.k.-modem for TDMA systems, International Conference on digital satellite communications (London), IEE Conference publication No. 59, November 1969.

10. LUNDQUIST, L., LAPRIORE, M. *and* GARDNER, F.M. 1974. Transmission of 4ϕ-phase-shift keyed time-division multiple access over satellite channels, IEEE Trans. Comms. 22, 9, September 1974.

11. COHN, S.B. 1957. Direct-coupled-resonator filters, Proc. IRE, 45, p 187, February 1957.

12. HARVEY, R.V. 1970. Field store standards conversion: equalisers for ultrasonic quartz delay lines, BBC Research Department Report No. 1971/36.

13. SUNDE, E.D. 1961. Pulse transmission by AM, FM and PM in the presence of phase distortion, BSTJ 40, 2, March 1961.

14. ROBINSON, R.N. 1971. An experimental quadrature modulation system for video signals, BBC Research Department Report No. 1971/30.

15. GARDNER, F.M. 1966. Phase-lock techniques. New York, Wiley, 1966.

16. BENNET, W.R. 1958. Statistics of Regenerative

Digital Data Transmission, BSTJ, 37, 6, November 1958.

17. REID, D.F. *and* CROLL, M.G. 1974. Digital sound signals: the effect of transmission errors in a near-instantaneous digitally companded system, BBC Research Department Report No. 1974/24.

18. BALL, J.R. 1975. Digital modems for microwave systems, paper presented at IEE colloquium, Digital microwave relay systems above 10 GH_z, April 1975.

19. MANSON, W.I. 1974. Digital sound signals; subjective effect of timing jitter, BBC Research Department Report No. 1974/11.

20. BENNET, W.R. 1960. Electrical Noise, McGraw-Hill, 1960.

APPENDIX

The loss of noise-margin due to phase jitter on the demodulating carrier

The following analysis shows that when the bit-error rate is 1 in 10^5 corresponding to a measured carrier-to-noise ratio of 13·4 dB, the phase jitter of 1·8° r.m.s. on the demodulating carrier degrades the noise margin of the modem by 0·1 dB.

Referring to Fig. 18, the vector OP of unit amplitude represents the d.p.s.k. signal occupying one of the four possible phase-states.

If the noise $\nu(t)$ added to the signal has an r.m.s. value σ the carrier-to-noise ratio

$$= 20 \log_{10} \frac{1}{\sigma\sqrt{2}} \text{ dB} = 13{\cdot}4 \text{ dB} \qquad \ldots (1)$$

The added noise $\nu(t)$, can be resolved into two quadrature components such that:

$$\nu(t) = x(t) \cos \omega_c t + y(t) \sin \omega_c t \qquad \ldots (2)$$

where ω_c = the carrier frequency in rads/s.

The quantities $x(t)$ and $y(t)$ are uncorrelated Gaussian variables and they also have the same r.m.s. value, σ as $\nu(t)$.[20] The quadrature noise components are shown as x(t) and y(t) on the vector diagram. The r.m.s. values of x(t) and y(t) also equal σ.

The dotted lines in Fig. 18, represent the decision axes of the synchronous demodulation process. The orientation of these axes will vary in sympathy with the phase jitter on the demodulating carrier. They can be restored to a stationary position on the vector diagram by adding a complementary phase jitter to OP. This is achieved by adding the vector shown as ϕ(t). We can write that:

the r.m.s. carrier phase jitter =

$$= \text{r.m.s. value of } \phi(t) \text{ rads/s} \qquad \ldots (3)$$

let the r.m.s. value of $\phi(t) = \epsilon$, therefore:

the r.m.s. carrier phase jitter =

$$= \frac{\epsilon \cdot 180}{\pi} \text{ degrees} = 1{\cdot}8^{\circ} \text{ (given above)} \qquad \ldots (4)$$

The vector ϕ(t) can in turn be resolved into two quadrature components with r.m.s. value $(\epsilon)/(\sqrt{2})$ in the same direction as the noise vectors x(t) and y(t). The effective value of the noise vectors is increased by the carrier phase jitter and consequently the noise margin of the modem is reduced. The effective r.m.s. value of the noise vectors x(t) and y(t) is given below.

$$\text{Effective r.m.s. value of x(t) and y(t)} = \left[\sigma^2 + \frac{\epsilon^2}{2}\right]^{1/2} \qquad \ldots (5)$$

The loss of noise margin of the modem due to phase jitter on the demodulating carrier is therefore

$$= 10 \log_{10} \frac{\left[\sigma^2 + \frac{\epsilon^2}{2}\right]}{\sigma^2} \text{ dB} \qquad \ldots (6)$$

By substituting for the values of σ and ϵ obtained by solving equations 1 and 4, the loss of noise margin was calculated as 0·1 dB.

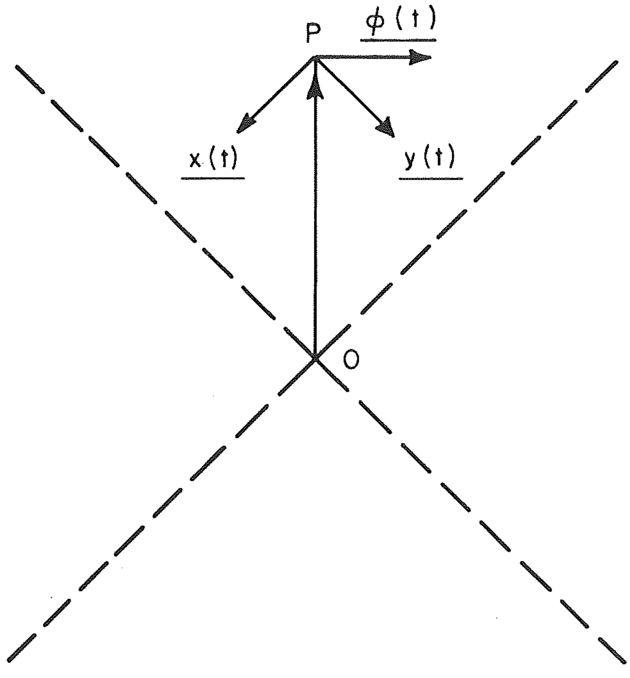


Fig. 18 - Vector diagram to illustrate the effect of phase jitter on the demodulating carrier

HR/JUC

Printed by BBC RESEARCH DEPARTMENT, Kingswood Warren, Tadworth, Surrey, KT20 6NP